Application of: McCallister, et al.	Date: 24 September 2005
Serial Number: 10/718,505	Group Art Unit: 2637
Filed: 20 November 2003	Examiner: CORRIELUS, JEAN B.
Title: *"Constrained-Envelope Transmitter and Method Therefor"*	Attorney Docket No.: 1826-310CIPRI

Assistant Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

APPELLANT'S BRIEF

Dear Sir:

This Brief is filed pursuant to a Notice of Appeal mailed on 24 May 2005 in the matter of the above-identified Reissue application.

Real Party in Interest

Intersil Americas, Inc. is the real party in interest and the assignee of this Reissue application.

Related Appeals and Interferences

Appellant is aware of no related appeals, interference, and/or other proceedings relevant to this discussion.

Status of Claims

Claims 1-28, of which claims 1, 12, 18, and 21 are independent claims, have been presented in this Reissue application. The Final Office Action rejected claims 1-28. Claims 2, 4-5, 14-16, and 18-28 have been cancelled by the appellants in an Amendment After Final submitted with this Appeal Brief. Accordingly, claims 1, 3, 6-13, and 17 are on appeal.

Appendix A provides a copy of all claims on appeal.

The Final Office Action rejected all claims on appeal under 35 U.S.C. 103(a) as obvious over May & Rohling, "Reducing the Peak-to-Average Power Ratio in OFDM Radio Transmission Systems," 48th IEEE Vehicular Technology Conference, 18-21 May 1998, 2474-2478 (hereinafter *May*) in view of Briffa et al., U.S. Patent No. 6,075,411 (hereinafter *Briffa*).

Appendix B provides copies of the *May* and *Briffa* references.

Status of Amendments

An Amendment After Final is being filed with this Appeal Brief to clarify arguments presented herein and remove issues for appeal. The Amendment cancels claims 2, 4-5, 14-16 and 18-28, and changes the dependency of claim 3 from canceled-claim 2 to pending-claim 1. This Amendment has not been acted upon by the Examiner.

Summary of Claimed Subject Matter

The present application is for a reissue of McCallister et al., U.S. Patent No. 6,366,619 (hereinafter *McCallister*).

There are no Figures relevant to the present discussion beyond those within *McCallister*.

The present invention pertains to constrained envelope digital transmitter circuits.

All references to column and line numbers in this Summary are taken from *McCallister*.

Transmitter circuit 22 (FIG. 2) has a binary data source 32 providing a binary input signal stream 34 of to-be-communicated data [column 4, lines 26-28]. Binary input signal stream 34 passes into a modulator 77, formed of a convolutional encoder 36, an interleaver 40, a phase mapper 44, and a pulse-spreading filter 76 [column 4, line 39, to column 7, line 10]. The output of modulator 77 is modulated signal 74, and conveys the data originally presented in binary input signal stream 34, albeit in a modulated form [column 7, lines 10-14].

In accordance with Shannon's theory, pulse-spreading filter 76 produces at least two output filtered-signal pulses 78 for each input phase-point pulse 66 received (FIG. 4) [column 7, lines 15-19]. In effect, modulated signal 74 is made up of two interleaved data streams, an on-time signal stream 84 and an off-time signal stream 86 [column 7, lines 26-28].

Off-time signal stream 86 passes from an output of modulator 77 to an input of an off-time constrained-envelope generator 106, which produces an off-time constrained-bandwidth error signal stream 108 from off-time signal stream 86 [column 9, lines 47-53]. Constrained-envelope signal stream 112 is effectively modulated signal 74 exhibiting a relatively low peak-to-average power ratio characteristic more easily accommodated by downstream components [column 9, lines 58-65].

On-time signal stream 84, also a portion of modulated signal 74, passes from an output of modulator 77 to an input of an on-time constrained-envelope generator 106', which produces an on-time constrained-bandwidth error signal stream 108' [column 12, lines 11-16]. Combining circuit 110 combines both off-time and on-time constrained-bandwidth error signal streams 108 and 108' with delayed modulated signal 140 to produce altered modulated signal 112 [column 12, lines 16-20].

A complex summing or combining circuit 110 combines off-time constrained-bandwidth error signal stream 108, on-time constrained-bandwidth error signal stream 108', and delayed modulated signal 140 to produce a constrained-envelope signal stream 112, which is an altered version of modulated signal 74 [column 9, lines 53-58].

Constrained-envelope signal 112 is passed to an input of a substantially linear amplifier 146 (FIG. 2) [column 14, lines 4-6]. Substantially linear amplifier 146 produces RF broadcast signal 26, which is then broadcast via transmitter antenna 24 [column 14, lines 6-8].

Substantially linear amplifier 146 is made up of a digital linearizer 148, a digital-to-analog converter 150, and a radio-frequency (RF) amplifying circuit 152 [column 14, lines 8-11]. Digital linearizer 148 alters constrained-envelope signal stream 112 into a pre-distorted digital signal stream 154, which is non-linear in just the right manner to compensate for non-linearities within digital-to-analog converter 150 and RF amplifying circuit 152, hence linearizing substantially linear amplifier 146 [column 14, lines 11-24].

Digital-to-analog converter 150 then converts pre-distorted digital signal stream 154 into an analog baseband signal 156, which is then amplified by RF amplifying circuit 152 into RF broadcast signal 26 and transmitted via transmitter antenna 24 [column 14, lines 25-29]. Due to the combination of constrained error signals 108 with modulated signal 74, substantially linear amplifier 146 need not process as great a peak-to-average power ratio as would be required if constrained error signals 108 were not used, while constrained error signals 108 are configured so as to substantially prevent the bandwidth of modulated signal 74 from increasing [column 14, lines 29-36].

Grounds of Rejection to Be Reviewed on Appeal

The following grounds of rejection are presented for review:

1. Whether independent claims 1 and 12 and claims dependent therefrom are made obvious under 35 U.S.C. 103(a) over *May* in view of *Briffa*.

Arguments

Grounds of Rejection 1 -- Claims 1, 3, 6-13, and 17

A Final Office Action dated 9 March 2005 rejected all claims on appeal under 35 U.S.C. 103(a) as being obvious over *May* in view of *Briffa*. Claims 1 and 12 are independent claims. The remainder of the claims on appeal depend directly or indirectly from either claim 1 or claim 12.

May proposes the mathematical basis for a peak-to-average power ratio (PAPR) reduction circuit in the context of orthogonal frequency division multiplex (OFDM) modulation. The Final Office Action asserts that *May* discloses those elements of the appellants' independent claims 1 and 12 that relate to a PAPR reduction circuit, but fails to discloses limitations directed to the use of linearization and/or power amplification. The final Office Action also asserts *Briffa* is in the same field of endeavor as *May* and teaches the use of linearization and power amplification. The Final Office Action alleges that it would have been obvious to one of ordinary skill in the art to combine the teachings of *Briffa* with those of *May* in order to adjust the amplitude and phase of the input signal.

Appellants respectfully disagree.

MPEP E8r2 70-6.02(j) states [emphases added]:

> To establish a prima facie case of obviousness, three basic criteria must be met. First, **there must be some suggestion** or motivation, either in the references themselves or in the knowledge generally available to one of ordinary skill in the art, to modify

> the reference or **to combine reference teachings**. Second, **there must be a reasonable expectation of success**. Finally, the prior art reference (or references when combined) must teach or suggest all the claim limitations. The teaching or suggestion to make the claimed combination and the reasonable expectation of success must both be found in the prior art and not based on applicant's disclosure.

In this situation, the prior art fails to provide a suggestion or motivation for combining references and further fails to provide a reasonable expectation of success. Accordingly, a prima facie case for obviousness fails, and the claims on appeal should be found allowable.

Briffa discloses a method and apparatus for the linearization of a signal to be amplified by a non-linear amplifier. That the RF amplifier of *Briffa* is nonlinear is made clear where *Briffa* discusses different linear and nonlinear classes of amplifiers in column 1, lines 23-33 and concludes that "because of their efficiency, nonlinear amplifiers are largely preferred, leaving the problem of distortion to deal with." The remainder of the *Briffa* disclosure is dedicated to dealing with the distortion problem of nonlinear amplifiers. In addition, *Briffa* in column 4, lines 23-42, discusses a problem associated with operating the amplifier in saturation, which is a characteristic of nonlinear amplifiers. Then, in column 5, line 4, *Briffa* explicitly says that the "predistortion signal compensates for IMD products produced by a nonlinear amplifier," and in column 7 line 22 *Briffa* says that "[s]ince the RF PA 13 has a complex gain which varies somewhat as the RF input level is varied i.e., it is nonlinear …"

The distinction between a linear and nonlinear amplifier is

no trivial matter in the context of RF transmitters. Appendix B provides evidence characterizing some of the significant differences between linear and nonlinear amplifiers as understood by those skilled in the art. For example, the tutorial provided in Appendix B entitled "Clases of Amplifier Operation" confirms in the last paragraph that Class C amplifiers "generally are operated in the region of plate- or collector-saturation …" and, the Power Amplifier chapter from the dissertation provided in Appendix B provides a tutorial of the various classes of amplifiers, and confirms, on page 13, that those skilled in the art understand class A, class B, and class AB amplifiers to be considered linear amplifiers while class C is considered a nonlinear class.

May provides a sparse disclosure that proposes an idea for the reduction of peak-to-average power ratio in a transmission system that uses a linear amplifier. That the RF amplifier contemplated for the proposal of *May* is linear is made clear by Fig. 1 and the references to Fig. 1. *May's* Fig. 1 shows a transfer curve for what is called an ideal limiter. In the last paragraph in the first column on page 2474 *May* indicates that for *May's* mathematical purposes the amplifier is to be modeled as an ideal limiter. In accordance with *May's* ideal limiter, the hypothetical and ideal amplifier/limiter is perfectly linear up to the point of saturation. In accordance with the *May* idea, as summarized in the second to last paragraph on page 2476, an input backoff is chosen so that most amplitude peaks presented to the input of the ideal limiter fall below the saturation point, and the few amplitude peaks that might otherwise exceed the saturation point are reduced by an additive correction signal *k(t)* so that no peaks actually exceed the saturation

point of the amplifier.

Those skilled in the art would not combine the *May* peak reduction circuit with the *Briffa* linearizer, as suggested by the Final Office Action, because *May* and *Briffa* are incompatible with each other. In particular, those skilled in the art would understand that no improvement would result to either *May* or *Briffa* since *Briffa* is intended to operate with a nonlinear amplifier that spends a significant amount of time in saturation and *May* is intended to work with an ideal linear amplifier that is prevented from ever going into saturation. Consequently, there is no motivation for making such a combination and no reasonable expectation of success to be achieved by such a combination. Independent claims 1 and 12, and the claims that depend therefrom should be found allowable over the combination of *May* and *Briffa*.

The conclusion that independent claims 1 and 12 should be found allowable over *May* in combination with *Briffa* may also be reached again from an entirely different direction. Those skilled in the art will appreciate that amplifiers are often classified into classes, such as A, B, AB, C and the like (see items 3 and 4 in Appendix B). Classes A, B, and AB are called linear amplifiers (see item 4 in Appendix B, page 13) in the art because they have high fidelity. In other words, they tend to faithfully reproduce their input signal at the output, only amplified. But this linearity comes at the cost of efficiency. Class C amplifiers are considered nonlinear (see item 4 in Appendix B, page 13) and are at the other end of the scale. They are the highly efficient, but achieve their great efficiency at the cost of poor fidelity. In other words, they

tend to produce a lot of distortion, both in-band and out-of-band. This much is also discussed in the background section in *Briffa*.

In some applications it is desirable to use a nonlinear, class C amplifier to take advantage of its efficiency, then use other techniques to improve its nonlinear nature to be more linear, as discussed in *Briffa*. Naturally, but misleadingly, such techniques may be called linearization. Clearly, such techniques do not make nonlinear amplifiers as linear as a linear amplifier. There would be no need -- ever -- for a linear amplifier if a much more efficient nonlinear amplifier could be made as linear as a linear amplifier. Rather, linearization in the context of nonlinear amplifiers simply make them more linear than they could be without linearization and in any event sufficiently linear to accommodate whatever specifications are imposed on the communication system. The out-of-band distortion generated in large quantity by nonlinear amplifiers may be ameliorated using techniques known to those skilled in the art, such as tank circuits, filters on output signals, and the linearization taught by *Briffa*. But these techniques, used alone or in combination with one another, merely mitigate the spectrum broadening effect of out-of-band distortion to a tolerable level for a given set of specifications.

May, on the other hand, is concerned with the entirely different problems imposed by using a linear amplifier. *May*, on page 2475, critiques a prior technique of correcting an OFDM signal by using a multiplicative correcting function. The use of a multiplicative correcting function is similar to the *Briffa*

linearization technique. In this passage *May* notes that with judicious selection of an attenuation function, spectrum broadening may be held to an acceptable level. Such a technique might very well be compatible with *Briffa* because the use of nonlinear amplification in *Briffa* broadens spectrum anyway.

But the idea proposed in *May* is indicated as producing no out-of-band interference (see *May*, page 2476, first column, sixth line), i.e., no spectrum broadening. Those skilled in the art would not combine *May* with *Briffa* because no improvement would result since *May* does not broaden the signal spectrum but *Briffa* permits a tolerable degree of spectrum broadening. All the good things that *May* accomplishes without causing interference in adjacent frequency bands are undone by *Briffa's* tolerable spectrum broadening. One skilled in the art would not make a combination that would add complication and achieve no benefit. Consequently, there is no motivation for making such a combination and no reasonable expectation of success to be achieved by such a combination. Independent claims 1 and 12, and the claims that depend therefrom should be found allowable over the combination of *May* and *Briffa*.

The conclusion that there is no suggestion to combine the sparsely disclosed *May* idea with the teaching of *Briffa* can be reached yet again from still another direction. In particular, if such a combination were made, the resulting performance would actually be worse than either *May* or *Briffa* alone contemplates.

May's proposed idea may be viewed as a transformer. It transforms the greatest signal peaks into in-band distortion (see 2476, first column, lines 6-7) without generating out-of-

band distortion. One of the problems facing a transmitter designer attempting to implement the *May* idea would be to use a threshold that minimally applies this transformation so that the in-band interference is at minimal power. But any in-band interference worsens performance, and the idea proposed in *May* transforms all peak energy above a threshold into nothing other than in-band distortion. The selective peak reduction that the *May* idea proposes, for a given communication signal scenario, allows the use of a linear amplifier having a lower saturation point to be used without causing nonlinear operation of the amplifier. In other words, the *May* idea allows a transmitter to operate at an average power level beneath but closer to the linear amplifier's saturation point without causing the amplifier to actually go into saturation.

Briffa, through the use of a nonlinear amplifier, is already designed to routinely operate at or above the amplifier's saturation point. Otherwise, the power efficiency benefits of a nonlinear amplifier are not achieved. Thus, *Briffa* cannot take advantage of *May's* minimal peak reduction. In fact, the purpose of *Briffa* is to reduce the predistortion that is applied when operating above the amplifier's saturation point because the predistortion signal during saturation tends to make the phase corrections "grossly incorrect" (see Briffa at column 4, line 39). Consequently, peak reduction is of no benefit to operation of the *Briffa* nonlinear amplifier.

With the proposed combination of the *May* idea feeding the *Briffa* linearizer and power amplifier, the result is *May's* in-band distortion and interference, including phase distortion, with no amplifier benefit in *Briffa*. This is worsened

performance. One skilled in the art would not be motivated to make a combination that resulted in worsened performance. Consequently, there is no motivation for making such a combination and no reasonable expectation of success to be achieved by such a combination. Independent claims 1 and 12, and the claims that depend therefrom should be found allowable over the combination of *May* and *Briffa*.

Accordingly, for the several reasons set forth above, independent claims 1 and 12 as well as the claims that depend therefrom are not obvious over the combination of *May* and *Briffa* and should be found allowable.

Conclusion

Claims 1, 3, 6-13, and 17 are included in this appeal. The rejection of the claims on appeal under 35 U.S.C. 103(a) as obvious over *May* in view of *Briffa* is believed to be improper.

Appellant believes that the arguments above fully respond to every outstanding ground of rejection and that the contested claims should be found allowable.

Respectfully submitted,

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Appendix A -- Claims on Appeal

This Appendix is seven pages, including this cover page, and contains a double-spaced copy of the claims on appeal.

In accordance with the guidelines for a reissue application, as put forth in MPEP 1452, the claims herein include all underlined and bracketed material necessary to reflect the changes made to the claims during the prosecution of the reissue application.

Claim 1: A constrained-envelope digital communications transmitter circuit comprising:

a modulated-signal generator for generating a first modulated signal conveying to-be-communicated data, having a first bandwidth and having a first peak-to-average amplitude ratio;

a constrained-envelope generator for generating a constrained bandwidth error signal in response to said first modulated signal;

a combining circuit for combining said constrained bandwidth error signal with said first modulated signal to produce a second modulated signal conveying said to-be-communicated data, said second modulated signal having substantially said first bandwidth and a second peak-to-average amplitude ratio, said second peak-to-average amplitude ratio being less than said first peak-to-average amplitude ratio; [and]

[a substantially linear amplifier configured to amplify said second modulated signal]

<u>a linearizer configured to pre-distort said second modulated signal into a pre-distorted signal; and</u>

<u>a radio-frequency amplifying circuit configured to generate a radio-frequency broadcast signal from said pre-distorted signal</u>.

Claim 2: Canceled.

Claim 3: A constrained-envelope digital communications transmitter circuit as claimed in claim [2] 1, wherein said constrained-envelope generator is configured so that said constrained bandwidth error signal exhibits a bandwidth substantially equal to or less than said first bandwidth.

Claims 4-5: (Canceled).

Claim 6: A constrained-envelope digital communications transmitter circuit as claimed in claim 1 wherein said modulated-signal generator is a code division multiple access (CDMA) modulator and said first modulated signal conveys a plurality of code-channels of said to-be-communicated data.

Claim 7: A constrained-envelope digital communications transmitter circuit as claimed in claim 6 wherein said CDMA modulator includes a Nyquist-type pulse spreading filter which provides said first modulated signal.

Claim 8: A constrained-envelope digital communications transmitter circuit as claimed in claim 1 wherein said constrained-envelope generator comprises:

a pulse generator responsive to said first modulated signal; and

a filter having an input coupled to said pulse generator and being configured to generate said constrained bandwidth error signal.

Claim 9: A constrained-envelope digital communications transmitter circuit as claimed in claim 8 wherein said pulse generator is configured to generate a pulse when said first modulated signal exhibits a magnitude greater than a threshold.

Claim 10: A constrained-envelope digital communications transmitter circuit as claimed in claim 9 wherein said pulse generator is further configured so that said pulse exhibits an amplitude which is responsive to a value by which said first modulated signal exhibits said magnitude greater than said threshold.

Claim 11: A constrained-envelope digital communications transmitter circuit as claimed in claim 1 wherein said [substantially linear amplifier comprises:]

[a linearizer configured to pre-distort said second modulated signal into a pre-distorted signal; and]

[a radio-frequency amplifying circuit configured to generate a radio-frequency broadcast signal from said pre-distorted signal]

<u>linearizer is a digital linearizer, and said transmitter circuit additionally comprises a digital-to-analog converter coupled between said digital linearizer and said radio-frequency amplifying circuit</u>.

Claim 12: In a digital communications system, a method for transmitting a constrained-envelope communications signal comprising:

generating a first modulated signal conveying to-be-communicated data and having a first bandwidth and a first peak-to-average amplitude ratio;

generating a constrained bandwidth error signal in response to said first modulated signal;

combining said constrained bandwidth error signal with said first modulated signal to produce a second modulated signal conveying said to-be-communicated data, said second modulated signal having substantially said first bandwidth and a second peak-to-average amplitude ratio, said second peak-to-average amplitude ratio being less than said first peak-to-average amplitude ratio; [and]

[linearly amplifying said second modulated signal]

linearizing said second modulated signal to produce a pre-distorted signal;

amplifying said pre-distorted signal to generate a communications signal exhibiting a constrained envelope; and

transmitting said communications signal.

Claim 13: A method as claimed in claim 12 wherein said constrained bandwidth error signal exhibits a bandwidth substantially equal to or less than said first bandwidth.

Claims 14-16: (Canceled).

Claim 17: A method as claimed in claim 12 wherein said first-modulated-signal-generating activity configures said first modulated signal as a code division multiple access (CDMA) signal conveying a plurality of code-channels of said to-be-communicated data.

Claims 18-28: (Cancelled).

Appendix B -- Evidence

This Appendix includes this cover page, and contains clean copies of all evidence (e.g., prior art references and the like) under consideration. This evidence is listed below:

1. May & Rohling, "Reducing the Peak-to-Average Power Ratio in OFDM Radio Transmission Systems," 48th IEEE Vehicular Technology Conference, 18-21 May 1998, 2474-2478 (5 pages).

	Patent	U.S. Pat. No.	Pages
2.	Briffa et al.	6,075,411	12

3. "Classes of Amplifier Operation", http://www.smeter.net/amplifiers/classes.php, 7/9/2005.

4. Saad Al-Shahrani, "Design of Class-E Radio Frequency Power Amplifier," Disertation submitted for the degree of Doctor of Philosophy in Electrical Engineering, Virginia Polytechnic Institute and State University, July 2001, Chapter 2 - Power Amplifier.

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Westin Hotel, Ottawa, Canada
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REDUCING THE PEAK-TO-AVERAGE POWER RATIO IN OFDM RADIO TRANSMISSION SYSTEMS

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Abstract—**An important difficulty which has to be solved in OFDM transmission systems is the large peak-to-average power ratio of the OFDM signal. Without any measures, the signal is limited by the power amplifier in the transmitter which causes interference both of the signal itself and in adjacent frequency bands. In this paper a method is proposed which considerably reduces the peak-to-average power ratio of the OFDM signal by means of signal processing.**

I. INTRODUCTION

OFDM systems allow the transmission of high data rates over broadband radio channels with frequency-selective fading without having to provide a powerful channel equalization. If differential modulation is applied, no channel estimation is required at all. For this reason, the complexity of OFDM systems can be much lower compared with a single carrier transmission system.

On the other hand, a difficulty about OFDM signals is the fact that they have a very large peak-to-average power ratio. In the transmitter, the maximal output power of the amplifier will limit the peak amplitude of the signal and this effect will produce interference both within the OFDM band and in adjacent frequency bands. Furthermore, in OFDM systems with a guard interval, a rectangular pulse must be used for modulation. The corresponding pulse spectrum is a sinc function which also causes out-of-band interference. These effects, particularly the problem of amplitude limitation of OFDM signals, have been discussed in several publications.

In this paper, we propose a manipulation of the OFDM signal like in [6] in order to remove peaks of the signal which exceed a given amplitude threshold. The proposed method does not produce out-of-band interference and with this condition it causes minimal interference within the OFDM band.

II. PROPOSALS IN THE LITERATURE

In most of the publications about amplitude limitation of OFDM signals it is assumed that it can be achieved by predistortion of the signal that the amplifier behaves like an ideal limiter. This means that the signal is amplified linearly up to a maximal input amplitude A_0 and larger amplitudes are limited to A_0, see Fig. . Based on this assumption, we also model the amplifier as an ideal limiter with amplitude threshold A_0 in this paper. The input power of the amplifier as compared to the threshold is described by the input backoff $IBO = 10\log\frac{A_0^2}{P_S}$ where P_S is the average OFDM signal power.

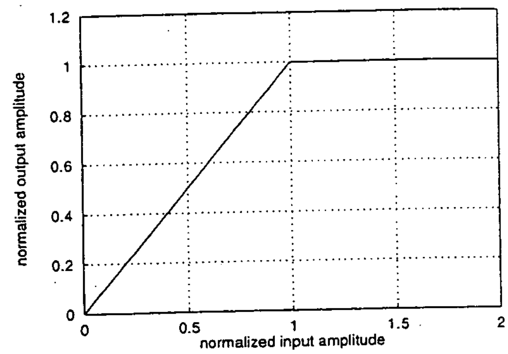

Fig. 1. Ideal limiter with normalized input and output amplitude, maximal input amplitude $A_0 = 1$

In literature, two kinds of approaches are investigated which assure that the transmitted OFDM signal $s(t)$ does not exceed the amplitude A_0 if a given input backoff is used. The first method makes use of redundancy in such a way that any data sequence leads to an OFDM signal with $|s(t)| \leq A_0$ or that at least the probability of higher amplitude peaks is greatly reduced. This approach does not result in interference of the OFDM signal.

In the second kind of approach, the OFDM signal is manipulated by a correcting function which eliminates the amplitude peaks. The out-of-band interference caused by the correcting function is zero or negligible. However, interference of the OFDM signal itself is tolerated to a certain extent. This approach will be investigated and optimized in the following.

A. Inserting Redundancy

One way to avoid high amplitude peaks is to select from the multitude of all possible OFDM blocks those which fulfill the condition $|s(t)| \leq A_0$ at a given input backoff. These "suitable" OFDM blocks could be assigned to different data bit sequences like a code. If e.g. only 10^{-9} of all possible blocks are selected, the system can transmit approximately 30 bits less per OFDM block, which could be acceptable. However, it seems to be necessary to find a way of constructing complying codes. If these codes cannot be constructed, the effort for this approach in terms of memory is generally too large. Even for a simple system with 16 subcarriers and QPSK, billions of assignments would have to be stored. In [2] short block codes are employed in order to enable a lower input backoff in OFDM systems with four or eight subcarriers. These codes can simultaneously be

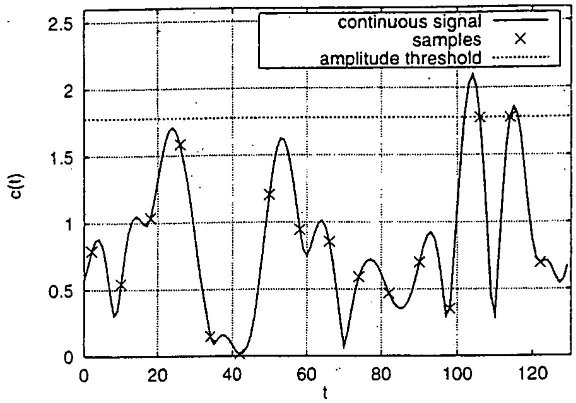

Fig. 2. The continuous-time OFDM signals exceeds the amplitude threshold A_0 in spite of clipping.

used for error correction.

In [4] procedures are proposed in which the same data sequence can be represented by several different OFDM blocks. The transmitter generates all possible signals corresponding to a data sequence and chooses the most suitable one for transmission. The receiver must additionally be told which of the signals has been chosen. This can be achieved with little redundancy. If differential modulation is applied between adjacent subcarriers, the receiver does not even need any side information. However, in this case on several subcarriers reference symbols are transmitted for the differential demodulation. This scheme allows e.g. to decrease the input backoff from 12 dB to 9 dB at the same level of out-of-band interference.

A proposal which realizes an OFDM transmission with a constant envelope using 50 % redundancy is proposed in [3]. In this scheme, instead of one OFDM block two blocks are transmitted which are calculated from $s(t)$. However, this calculation is non-linear and causes itself out-of-band interference. The objective of this approach is not to avoid out-of-band interference but to avoid interference of the OFDM signal.

B. *Correcting the OFDM – Signal*

The second approach in which the OFDM signal is corrected with a suitable function avoids out-of-band interference but it tolerates interference of the OFDM signal itself. In the simplest case, the sampled signal is limited to the amplitude A_0 [1]. This method is termed clipping. Clipping does not cause out-of-band interference if $s(t)$ is not oversampled. However, without oversampling the analog signal after the D/A conversion will exceed the amplitude threshold, see Fig. 2

This effect has to be considered. Furthermore, the OFDM signal must be filtered because of the rectangular pulse. For both reasons, oversampling of the signal is necessary. In [5] the proposal is made to apply clipping to the oversampled signal which causes out-of-band interference. This is taken care of by a FIR filter which also removes the side lobes of the modulation pulse. However, the filter leads to new amplitude peaks in the signal, but, after all, the peak-to-average power ratio of the signal is reduced by this method.

In [6], the OFDM signal is corrected by multiplying it with a correcting function $k(t)$. If the signal exceeds the amplitude threshold A_0 at the times t_n, then the corrected signal $c(t)$ is

$$\begin{aligned} c(t) &= s(t)k(t) \quad \text{where} \\ k(t) &= 1 - \sum_n A_n g(t - t_n) \\ g(t) &= \mathrm{e}^{-t^2/2\sigma^2} \\ A_n &= \frac{|s(t_n)| - A_0}{|s(t_n)|} \end{aligned}$$

Thus, the signal is attenuated by a Gaussian function at all positions where it has high amplitude peaks. The spectrum $C(f)$ of the corrected signal is

$$\begin{aligned} C(f) &= S(f) * K(f) \\ &= S(f) - S(f) * \left[\mathrm{e}^{-f^2/2\sigma_f^2} \sum_n B_n \mathrm{e}^{\mathrm{j}2\pi t_n f} \right] \end{aligned}$$

where B_n are constants and σ_f^2 is the variance of the Gaussian function $g(t)$ in the frequency domain. So, the correction broadens the signal spectrum by the width of a Gaussian function with the variance $\sigma_f^2 = 1/2\pi\sigma^2$ The Gaussian function which is used for the correction should be narrow in the time domain so that only a small interval of the signal is attenuated. It should also be narrow in the frequency domain so that the signal spectrum is broadened as little as possible. If for an OFDM system with the bandwidth B a Gaussian function with $\sigma = 5/B$ is chosen, then $\sigma_f = B/10\pi$, which results in an acceptable broadening of the signal spectrum.

Obviously, with this method it can be achieved to limit the amplitude of the oversampled signal without causing out-of-band interference, except in a narrow frequency band adjacent to the OFDM band. However, if many amplitude peaks have to be corrected, the entire signal is attenuated and the peak-to-average power ratio of the signal cannot be improved beyond a certain figure.

This scheme of correcting the OFDM signal can be realized for any number of subcarriers and it does not need any redundancy. It causes interference of the OFDM signal, but this is of secondary importance in a fading environment in which OFDM is typically applied. The important task is to avoid out-of-band interference.

III. Additive Correcting Functions

Each manipulation of a signal can be seen as an additive correction. The corrected signal can be written as

$$\begin{aligned} c(t) &= s(t) + k(t) \quad \text{where} \\ k(t) &= \sum_n A_n g(t - t_n) \\ A_n &= -\left(|s(t_n)| - A_0\right) \frac{s(t_n)}{|s(t_n)|} \end{aligned}$$

The correcting function $k(t)$ composed with the auxiliary function $g(t)$ which must be normalized so that $g(0) = 1$. This

correction limits the signal $s(t)$ to A_0 at the positions t_n of amplitude peaks. Of course, the correction could cause the signal to exceed the amplitude threshold at a different position. We ignore this effect now and we will see that it is negligible later. In the following, we determine an auxiliary function $g(t)$ which produces no out-of-band interference and causes interference of the OFDM signal with minimal power.

For the sake of simplicity, only a single OFDM modulation interval with the extent $T_N = 1/\Delta f$ is considered which is repeated periodically. The auxiliary function is also periodical, then. In this case, the spectra of both the signal and the auxiliary function are discrete. Thus, an auxiliary function with a spectrum limited to the OFDM band can be entirely described in the frequency domain by its samples at the subcarrier frequencies.

$$g(t) = \sum_{k=0}^{N-1} G_k e^{j2\pi k\Delta f t} \quad (1)$$

$$g(0) = \sum_{k=0}^{N-1} G_k = 1 \quad (2)$$

The interference power is minimized, if we minimize the power P_g of $g(t)$.

$$P_g = \frac{1}{T}\int_0^T |g(t)|^2 dt = \sum_{k=0}^{N-1} |G_k|^2$$

If an auxiliary function $g(t)$ is chosen such that the phases of the coefficients G_k are not all equal, then a different function with lower power could be found in the following way: The power of the function $g'(t)$ with the coefficients $G'_k = |G_k|$ has the same power as $g(t)$ and $g'(0) > g(0) = 1$. If $g'(t)$ is normalized according to Eqn. (2), the result is an auxiliary function with lower power than $g(t)$. Therefore, the auxiliary function with minimal power has coefficients G_k which are all equal. We obtain

$$P_g = \sum_{k=0}^{N-1} G_k^2$$

Taking Eqn. (2) into account, P_g is minimal, if we chose $G_k = 1/N$ for all coefficients. Then, Eqn. (1) yields

$$\begin{aligned} g(t) &= \frac{1}{N}\sum_{k=0}^{N-1} e^{j2\pi k\Delta f t} && (3)\\ &= \sum_{n=-\infty}^{\infty} \mathrm{sinc}[\pi B(t - nT_N)]e^{j\pi B(t-T_N)} && (4)\\ &\approx \mathrm{sinc}(\pi B t)e^{j\pi B t} && (5)\end{aligned}$$

This approximation is valid for $t \in [-T_N/2, T_N/2]$. The sinc function is repeated periodically because only a single modulation interval is considered which is repeated periodically, too, leading to a discrete spectrum. If the entire OFDM signal is considered and not only a single modulation interval,

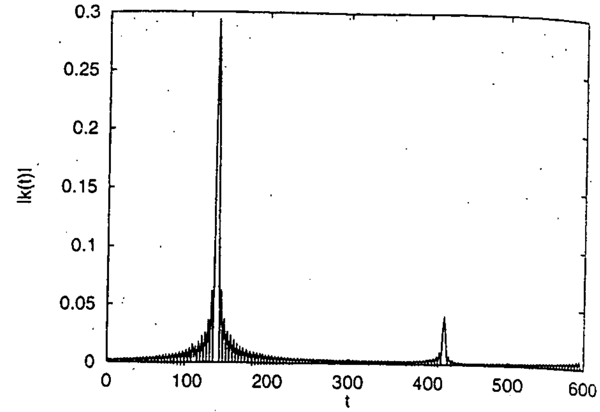

Fig. 3. Example of a correcting function $k(t)$ for an OFDM signal with 128 subcarriers, oversampling

then the auxiliary function is not periodical and a single sinc function with a continuous spectrum is obtained as it is given in Eqn. (5).

This function can correct an amplitude peak in the OFDM signals with minimal interference of the signal and without any out-of-band interference. For practical application however, the extent of the sinc function in time must be limited by windowing.

If the OFDM signal is not oversampled, then the sampled auxiliary function $g(n\Delta t)$ is zero for all $n \neq 0$. The correction scheme is identical with clipping in this case.

The spectrum of the corrected signal is

$$\begin{aligned} C(f) &= S(f) + K(f)\\ &= S(f) + \sum_n A_n G(f) e^{j2\pi t_n f}\\ &= S(f) + \sum_n \left[\frac{A_n}{N} e^{j2\pi t_n f} \sum_{k=0}^{N-1} \delta(f - k\Delta f)\right]\end{aligned}$$

Each correction of an amplitude peak causes interference on each subcarrier and the power of the correcting function is distributed evenly to all subcarriers.

This correcting scheme has been applied to an OFDM transmission system with 128 subcarriers in a simulation. The signal $s(t)$ has been oversampled by a factor of four and normalized so that the signal power is one. Then the signal has been corrected with $k(t)$. For the correction, the amplitude threshold A_0 has been set according to the input backoff which has been chosen. After the correction, the signal has been limited to the amplitude A_0 in order to take into account the limitation of amplitude peaks which may have remained.

Fig. 3 shows an example of a correcting function. In Fig. 4, the power density spectrum of the interference caused by the correction and the limitation of the signal is shown for an input backoff $IBO = 4$ dB. It can be seen that the interference power is concentrated within the OFDM bandwidth. Despite the correction of the OFDM signal, the signal still happens to exceed the amplitude threshold A_0. For this reason, the limiter causes

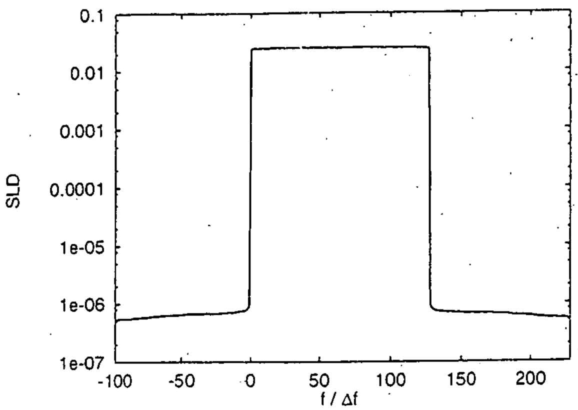

Fig. 4. Power density spectrum of the interference of the received signal which is caused by correction and limitation at $IBO = 4$ dB

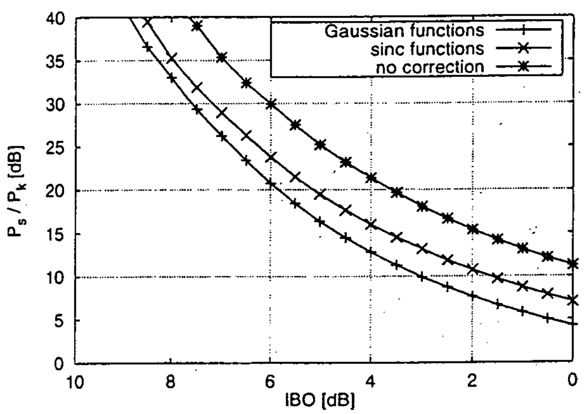

Fig. 5. Signal-to-interference ratio SIR in the OFDM band in the case of correction with Gaussian functions (multiplicative), with sinc functions (additive) and without any correction

out-of-band interference the power of which is more than 60 dB below the signal power.

The signal-to-interference power ratio SIR can be determined both within the OFDM band and in the adjacent frequency bands. The SIR in the OFDM band is the quotient of the power densities of the signal and of the interference. As the SIR in the adjacent frequency band we denote the quotient of the signal power density in the OFDM band and the average power density of the interference in a frequency band with the bandwidth $B/10$, directly adjacent to the OFDM band.

In Fig. 5 the resulting SIR in the OFDM band is depicted for correction with Gaussian functions (multiplicative), with sinc functions (additive) and without any correction. In each case, the signal has additionally been limited according to the IBO. Without any correction we obtain the least interference power, because the signal is modified only where the amplitude threshold of the limiter is exceeded. The correcting functions additionally modify the signal within a certain area around the amplitude peaks.

The SIR in the adjacent frequency bands is shown in Fig. 6.

The correction of the signal has the effect of an attenuation.

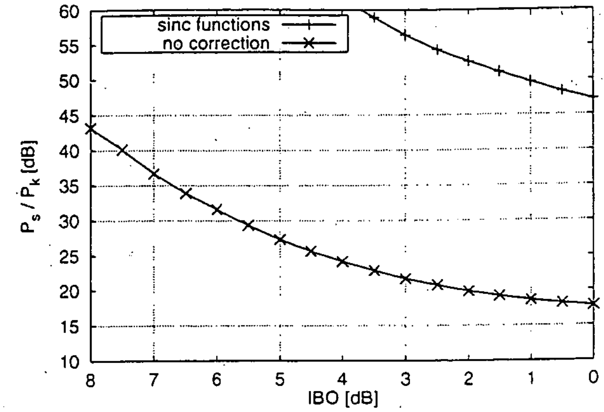

Fig. 6. Signal-to-interference ratio SIR adjacent to the OFDM band in the case of correction with sinc functions and without any correction

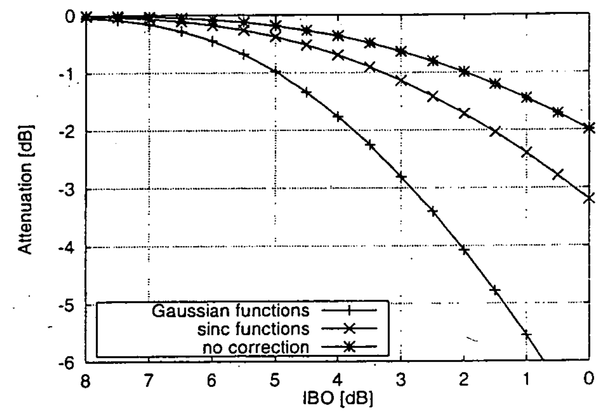

Fig. 7. Attenuation of the signal due to the correction with Gaussian functions, with sinc functions and with limitation of the signal

This is obvious in the case of the multiplicative correction with Gaussian functions, but the additive correction with sinc functions and the limitation of the signal also reduce the signal power. The effective attenuation of the signal due to the correction and the limitation is displayed in Fig. 7

Because of this attenuation of the signal, the real input power of the amplifier is lower than the IBO suggests which the correction of the signal was designed for. Note that IBO refers to the OFDM signal before the correction. The correction with sinc functions causes a lower attenuation than the correction with Gaussian functions so that the same IBO leads to a higher input power of the amplifier in this case.

For the AWGN channel an d DQPSK modulation, Fig. 8 shows the bit error rate as a function of the input backoff IBO. In the corresponding simulation, the noise power N has been fixed so that $A_0^2/N = 18$ dB. Thus, the signal-to-noise ratio SNR depends on the IBO. Still, the input backoff refers to the OFDM signal before correction.

For a WSSUS radio channel with multipath propagation and consequently with frequency-selective fading, Fig. 9 shows simulation results with $A_0^2/N = 30$ dB. In this case, the interference due to the correction is less grave, because it is attenuated by the radio channel as well, whereas the noise is added

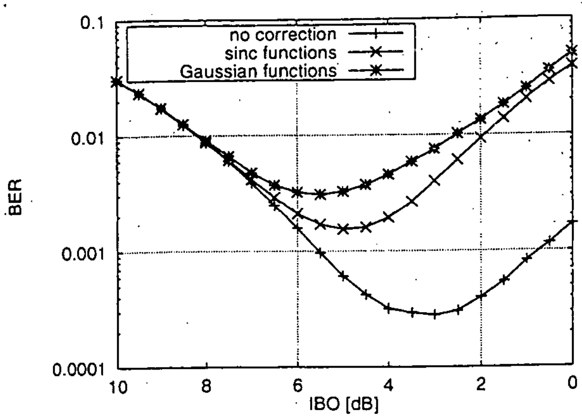

Fig. 8. Bit error rate BER as a function of the input backoff IBO of the limiter (amplitude threshold A_0) for the AWGN channel with $A_0^2/N = 18$ dB

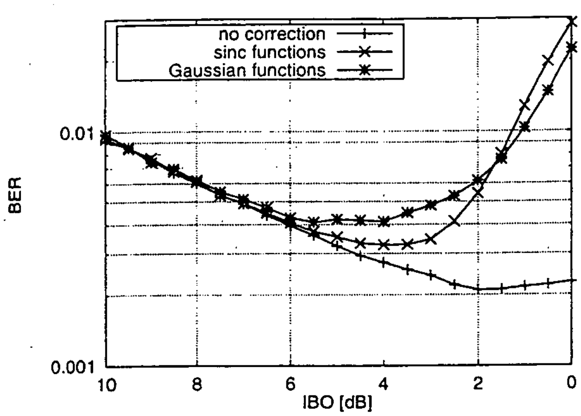

Fig. 9. Bit error rate BER as a function of the input backoff IBO of the limiter (amplitude threshold A_0) for a channel with frequency-selective Rayleigh fading and $A_0^2/N = 30$ dB

to the attenuated signal. Accordingly, a lower input backoff is optimum than in the case of the AWGN channel.

IV. Conclusion

In this paper the problem of out-of-band interference in OFDM transmission systems has been discussed which result from limiting the signal amplitude. The approach of correcting the OFDM signal with a suitable function has been analyzed. In this approach the signal is modified in such a way that a given amplitude threshold of the signal is not exceeded after the correction.

With the restriction that the correcting function does not cause any out-of-band interference, the correcting function has been given which has minimal power and thus causes minimal interference power within the OFDM band.

The interference power which is produced by the correcting function has been determined for the proposed method and for a multiplicative correction with Gaussian functions. Both for the AWGN channel and for a channel with frequency-selective fading with a fixed noise power the resulting bit error rate has been evaluated as a function of the input backoff of the power amplifier. It has been shown that for DQPSK modulation an input backoff of only 4 dB is reasonable with the proposed scheme.

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Classes of Amplifier Operation

RF amplifiers are classified A, AB, B or C according to the phase-angle (*number of degrees of current flow during each 360-degree RF cycle*) over which plate- or collector-current flows.

- **Class A Amplifiers**
- **Class B Amplifiers**
- **Class AB Amplifiers**
- **Class C Amplifiers**
- **Free Amplifier Design Software**

Class A Amplifiers
Class A amplifiers operate over a relatively small portion of a tube's plate-current or a transistor's collector-current range and have continuous plate- or collector-current flow throughout each RF cycle. Their efficiency in converting DC-source-power to RF-output-power is poor. DC source power that is not converted to radio frequency output power is dissipated as heat. However, in compensation, Class A amplifiers have greater input-to-output waveform linearity (*lower output-signal distortion*) than any other amplifier class. They are most commonly used in small-signal applications where linearity is more important than power efficiency, but also are sometimes used in large-signal applications where the need for extraordinarily high linearity outweighs cost and heat disadvantages associated with poor power efficiency.

Class B Amplifiers
Class B amplifiers have their tube control-grids or transistor bases biased near plate- or collector-current cutoff, causing plate- or collector-current to flow only during approximately 180 degrees of each RF cycle. That causes the DC-source-power to RF-output-power efficiency to be much higher than with Class A amplifiers, but at the cost of severe output cycle waveform distortion. That waveform distortion is greatly reduced in practical designs by using relatively high-Q resonant output "tank" circuits to reconstruct full RF cycles.

The effect is the same in principle as pushing a child in a swing through half-swing-cycles and letting the natural oscillatory characteristics of the swing move the child through the other half-cycles. However, low sine-wave distortion results in either case only if the Q of the oscillatory circuit (*the tank circuit or the swing*) is sufficiently high. Unless the Q is infinite, which it never can be, the amplitude of one-half cycle will be larger than the other, which is another way of saying there always will be some amount of harmonic energy. (*Coupling an antenna system too tightly to the resonant output tank circuit of an amplifier will lower its Q, increasing the percentage of harmonic content in the output.*)

Another effective method commonly used to greatly reduce Class B RF amplifier output waveform distortion (*harmonic content*) is to employ two amplifiers operating in "push-pull" such that one conducts on half-cycles where the other is in plate- or collector-current cutoff. Oscillatory tank circuits are still used in the outputs of Class B push-pull amplifiers to smooth switching transitions from the conduction of one amplifier to the other, and to correct other nonlinearities, but lower-Q tank circuits can be used for given percentages of harmonic content in the output. (*Tank circuits can be loaded more-heavily for given percentages of harmonic output where two amplifiers operate in push-pull.*)

Class AB Amplifiers
As the designation suggests, Class AB amplifiers are compromises between Class A and Class B operation. They are biased so plate- or collector-current flows less than 360 degrees, but more than 180 degrees, of each RF cycle. Any bias-point between those limits can be used, which provides a continuous selection-range extending from low-distortion, low-

efficiency on one end to higher-distortion, higher-efficiency on the other.

Class AB amplifiers are widely used in SSB linear amplifier applications where low-distortion and high power-efficiency tend to both be very important. Push-pull Class AB amplifiers are especially attractive in SSB linear amplifier applications, because the greater linearity resulting from having one amplifier or the other always conducting makes it possible to bias push-pull Class AB amplifiers closer to the Class B end of the AB scale where the power-efficiency is higher. Alternatively, push-pull Class AB amplifiers can be biased far enough toward the highly-linear Class A end of the scale to make broadband operation without resonant tank circuits possible in applications where broadband operation or freedom from tuning is more important than power-efficiency.

Class C Amplifiers
Class C amplifiers are biased well beyond cutoff, so that plate- or collector-current flows less than 180 degrees of each RF cycle. That provides even higher power-efficiency than Class B operation, but with the penalty of even higher input-to-output nonlinearity, making use of relatively high-Q resonant output tank circuits to restore complete RF sine-wave cycles essential. High amplifying-nonlinearity makes them unsuitable to amplify AM, DSB, or SSB signals.

However, most Class C amplifiers can be amplitude-modulated with acceptably low distortion by varying plate- or collector-voltage, because they generally are operated in the region of plate- or collector-saturation so that the RF output voltage is very closely dependent upon instantaneous DC plate- or collector-voltage. They also are commonly used in CW and frequency-shift-keyed radiotelegraph applications and in phase- and frequency-modulated transmitter applications where signal amplitudes remain constant.

DESIGN OF CLASS-E RADIO FREQUENCY POWER AMPLIFIER

by

Saad Al-Shahrani

Dissertation submitted to the Faculty of the
Virginia Polytechnic Institute and State University
in partial fulfillment of the requirements for the degree of

DOCTOR OF PHILOSOPHY

in

Electrical Engineering

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July 2001
Blacksburg, Virginia

DESIGN OF CLASS-E RADIO FREQUENCY POWER AMPLIFIER

by

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ABSTRACT

Power amplifiers (PA) are typically the most power-consuming building blocks of RF transceivers. Therefore, the design of a high-efficiency radio frequency power amplifier is the most obvious solution to overcoming the battery lifetime limitation in the portable communication systems. A power amplifier's classes (A, AB, B, C, F, E, etc), and design techniques (Load-pull and large-signal S-parameters techniques) are presented. The design accuracy of class-A power amplifier based on the small-signal S-parameters was investigated, where compression in the power gain was used as an indicator for design accuracy. The effect of drain voltage variation on the power gain compression has been studied in this research.

The class-E amplifier has a maximum theoretical efficiency of 100%. It consists of a single transistor that is driven as a switch and a passive load network. The passive load network is designed to minimize drain (collector) voltage and current waveforms overlapping, which minimize the output power dissipation. Two L-band class-E amplifiers are implemented in section 5.3. One of them is a lumped elements based circuit and the other is a transmission lines based circuit. Both circuits show good performance (60% PAD) over a wide bandwidth (1.0 GHz). In section 5.4, lumped elements and transmission lines based X-band class-E amplifiers are presented. Both circuits show good performance (62% PAD) over wide bandwidth (4.8 GHz).

A new technique to improve the drain efficiency of the class-E amplifier has been proposed. This technique uses two passive networks. One of them is in a series with the shunt capacitor C_S and the other is in a series with the transistor's source terminal. This technique shows improvement in the drain efficiency, which jumps from 62% to 82%.

Last few years have seen an increase in the popularity of the wireless communication systems. As a result, the demand for compact, low-cost, and low power portable (Single-chip) transceivers has increased dramatically. Among the transceiver's building blocks is the power amplifier. Thus, there is a need for a low-cost power amplifier. A 900 MHz CMOS RF PA with one-watt output power and a high power added efficiency (68%) is presented in chapter 6. This PA can be used in the European standard for mobile communications (GSM) handset transmitter.

AKNOWLEDGMENT

There are many people whom I would like to acknowledge for their assistance and support in completing this work, both technical and moral.

First, I would like to thank my advisor Dr. Sedki M. Riad, for his patient guidance and his generous support and encouragement during the course of my doctoral studies.

Appreciation is also extended to the members of my advisory committee, Drs. Ioannis Besieris, Aicha Elshabini, Amin Ezzeddine, and Jianqing He for serving on my committee as well as their invaluable guidance.

I would like also to acknowledge my family for their understanding and constant support during my study years.

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CHAPTER 2

POWER AMPLIFIER

2.0 Introduction

The main characteristics of an amplifier are Linearity, efficiency, output power, and signal gain. In general, there is a trade off between these characteristics. For example, improving amplifier's linearity will degrade its efficiency. Therefore knowing the importance degree of each one of these characteristics is an essential step in designing an Amplifier. This can be jugged based on the application. As an example high output power Amplifier is used in the transmitter side of a transceiver, whereas high linear amplifier used in the receiver side.

An amplifier is said to be linear if it preserves the details of the signal waveform, that is to say,

$$V_o(t) = A \cdot V_i(t) \tag{2.1}$$

where, V_i and V_o are the input and output signals respectively, and A is a constant gain representing the amplifier gain. But if the relationship between V_i and V_o contains the higher power of V_i, then the amplifier produces nonlinear distortion.

The amplifier's efficiency is a measure of its ability to convert the dc power of the supply into the signal power delivered to the load. The definition of the efficiency can be represented in an equation form as

$$\eta = \frac{Signal\ power\ delivered\ to\ load}{DC\ power\ Supplied\ to\ output\ circuit} \tag{2.2}$$

For an ideal amplifier, the efficiency is one. Thus, the power delivered to the load is equal to the power taken from the DC supply. In this case, no power would be consumed in the amplifier. In reality, this is not possible, especially in high frequency realm of RF circuits. In many high frequency systems, the output stage and driver stage of an amplifier consumed power in the amplification process.

The gain of the amplifier (G) is equal to the magnitude of the output signal (X_o) over the magnitude of the input signal (X_i) as shown in the equation.

$$G = \frac{X_o}{X_i} \quad . \tag{2.3}$$

G can be voltage, current, or power gain depending on the application.

The output power level plays an important role in evaluating the power amplifier. The power output capability factor, P_{MAX}, is the power output that would be produced with stresses of 1 Volt and 1 Amp on the drain of the field effect transistor (FET). Multiplication of P_{MAX} by the drain voltage and current ratings of a real device produces the maximum output power available from that device.

The power output capability factor is

$$P_{MAX} = \frac{\text{The Maximum Output Power}}{\text{The Peak Drain Voltage} \times \text{The Peak Drain Current}} \quad . \tag{2.4}$$

2.1 Amplifier Classification

Amplifiers are classified according to their circuit configurations and methods of operation into different classes such as A, B, C, and F. These classes range from entirely linear with low efficiency to entirely non-linear with high efficiency. The analysis presented in this chapter assumes piecewise-linear operation of the active device. The majority of this

information is available in Solid State Radio Engineering by Krauss, Bostain, and Raab [1980].

The active device used in this research is the field effect transistor. The reason for choosing this type of transistor is its superior performance in the microwave range The characteristics of the FET can be described by:

$$i_D = 0 \qquad \text{cut-off region,}$$

$$i_D = g_m \cdot (V_{GS} - V_T) \qquad \text{active region,} \tag{2.5}$$

$$i_D = \frac{V_D}{R_{on}} \qquad \text{saturation region.}$$

The regions of operation are defined by:

cut-off region: $V_{GS} < V_T$,

active region : $V_{GS} \geq V_T$ and $i_D < V_D/R_{on}$,

saturation region: $V_{GS} \geq V_T$ and $i_D = V_D/R_{on}$.

The term "saturation" is used here to denote the region where further increase in gate voltage produces no increase in drain current, that is to say, i_D is independent of V_{GS}.

2.2.1 Class A

The class-A amplifier has the highest linearity over the other classes. It operates in a linear portion of its characteristic; it is equivalent to a current source. As shown in figures.2.1 and 2.2, the configurations of class-A, B, and C amplifiers can be either a push–pull or a single ended tuned version. Figure.2.3 shows the load-line and current waveform for the class-A amplifier. To achieve high linearity and gain, the amplifier's base and drain dc voltage should by chosen properly so that the amplifier operates in the linear region. The device, since it is on

(conducting) at all times, is constantly carrying current, which represents a continuous loss of power in the device.

As shown in Fig.2.3, the maximum ac output voltage V_{om} is slightly less than V_{DD} and the maximum ac output current I_{om} is equal to I_{dq}. In the inductor-less system, the output voltage V_{om} will not be able to rise above the supply voltage, therefore, the swing will be constrained to $V_{DD}/2$ and not V_{DD}. The drain voltage must have a dc component equal to that of the supply voltage and a fundamental-frequency component equal to that of the output voltage; hence

$$V_D(\theta) = V_{DD} + V_{om} \cdot \sin\theta \quad . \tag{2.6}$$

The dc power is

$$P_{dc} = V_{DD} \cdot I_{dq} \quad , \tag{2.7}$$

the maximum output power is

$$P_o = \frac{1}{2} \cdot V_{om} \cdot I_{om} \approx \frac{1}{2} \cdot V_{DD} \cdot I_{dq} \quad , \tag{2.8}$$

and the efficiency is

$$\eta = \frac{P_o}{P_{dc}} \cdot 100 = \frac{1}{2} \cdot \frac{V_{om}}{V_{DD}} \cdot 100 \le 50\% \quad . \tag{2.9}$$

The difference between the dc power and output power is called power dissipation:

$$P_d = P_{dc} - P_o \quad . \tag{2.10}$$

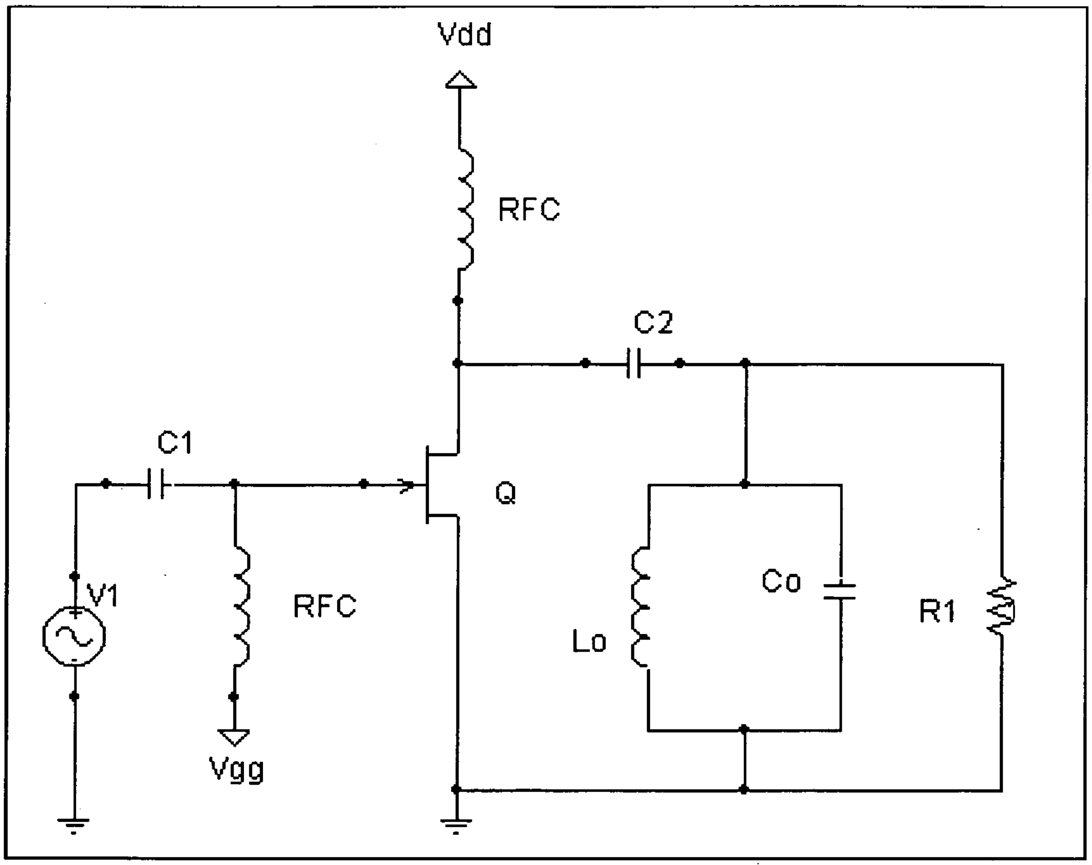
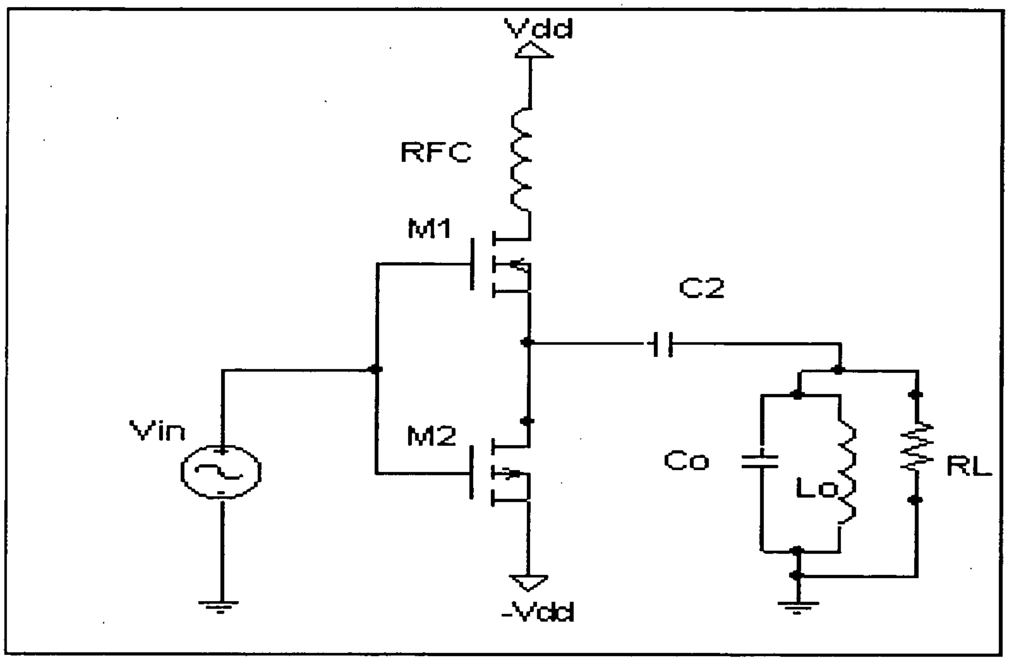
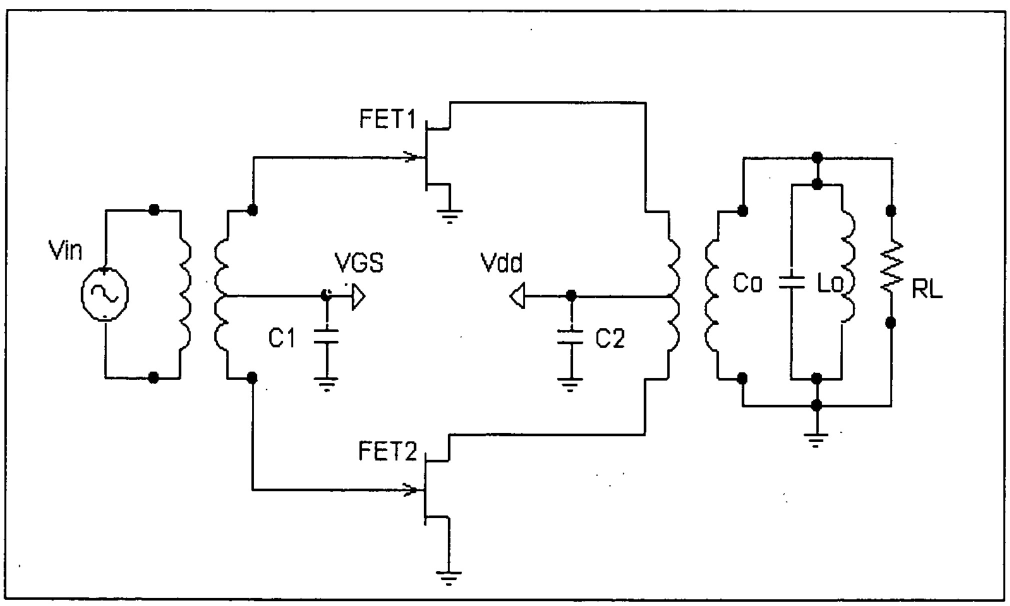


Figure 2.1. **Single-ended Power Amplifier (Class A, B, or C)**



(a)



(b)

Figure 2.2.

a. **Complementary Push–pull Power Amplifier (Class A, B, or C)**

b. **Transformer-coupled Push–pull Power Amplifier (Class A, B, or C)**

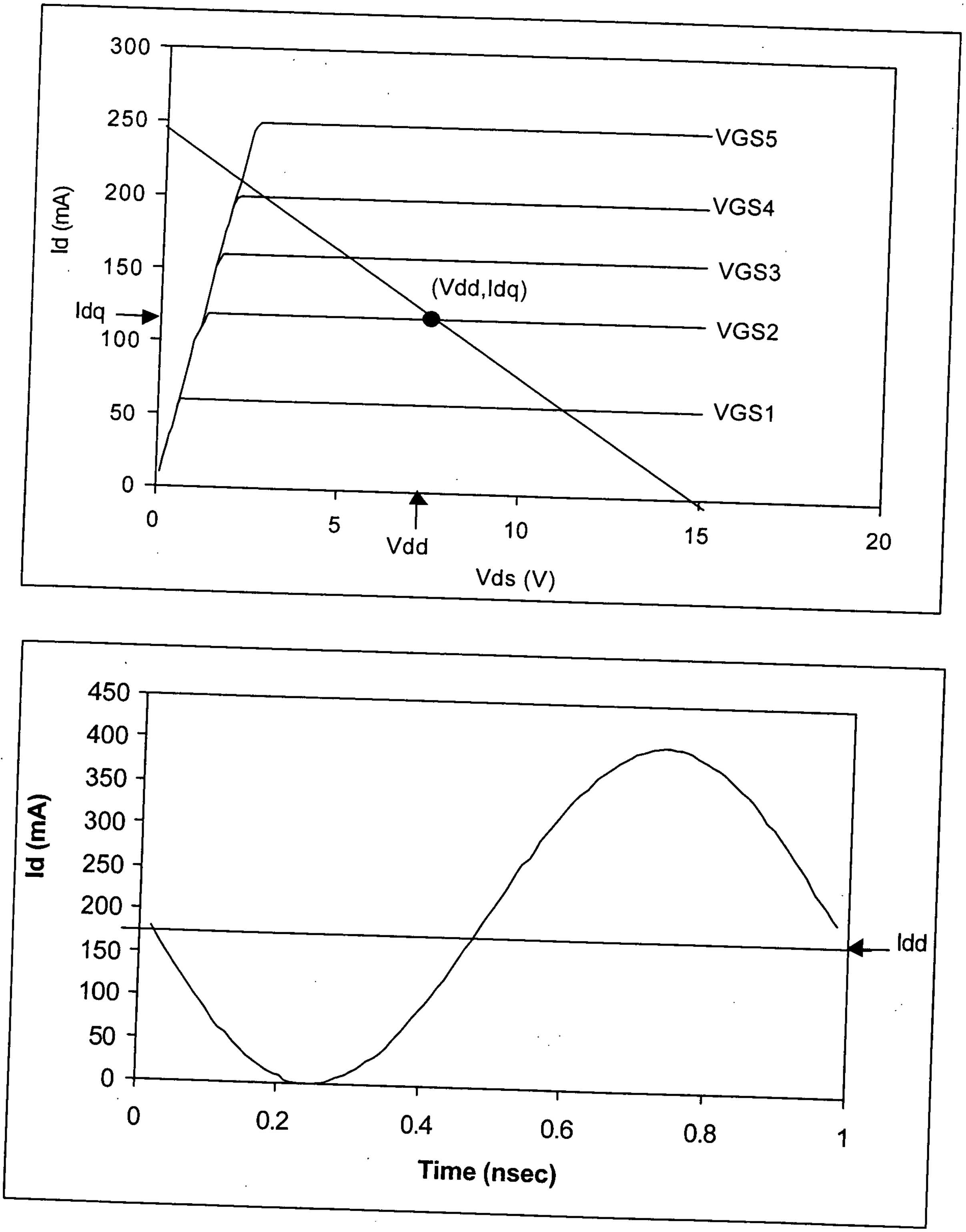


Figure 2.3. **Load line and current waveform for the class-A power amplifier**

2.2.2 Class B

The class-B amplifier operates ideally at zero quiescent current, so that the dc power is small. Therefore, its efficiency is higher than that of the class-A amplifier. The price paid for the enhancement in the efficiency is in the linearity of the device.

Figure 2.4 shows how the class-B amplifier operates. The output power for the single-ended class-B amplifier is

$$P_o = \frac{1}{2} \cdot I_{om} \cdot V_o \quad . \tag{2.11}$$

the dc drain current is

$$I_{dc} = 2\frac{I_{om}}{\pi} \quad , \tag{2.12}$$

the dc power is

$$P_{dc} = 2\frac{I_{om} \cdot V_{DD}}{\pi} \quad , \tag{2.13}$$

and the maximum efficiency when $V_{om} = V_{DD}$ is

$$\eta = \frac{P_o}{P_{dc}} \cdot 100 = \frac{\pi}{4} \cdot \frac{V_{om}}{V_{DD}} \cdot 100 \leq 78.53\% \quad . \tag{2.14}$$

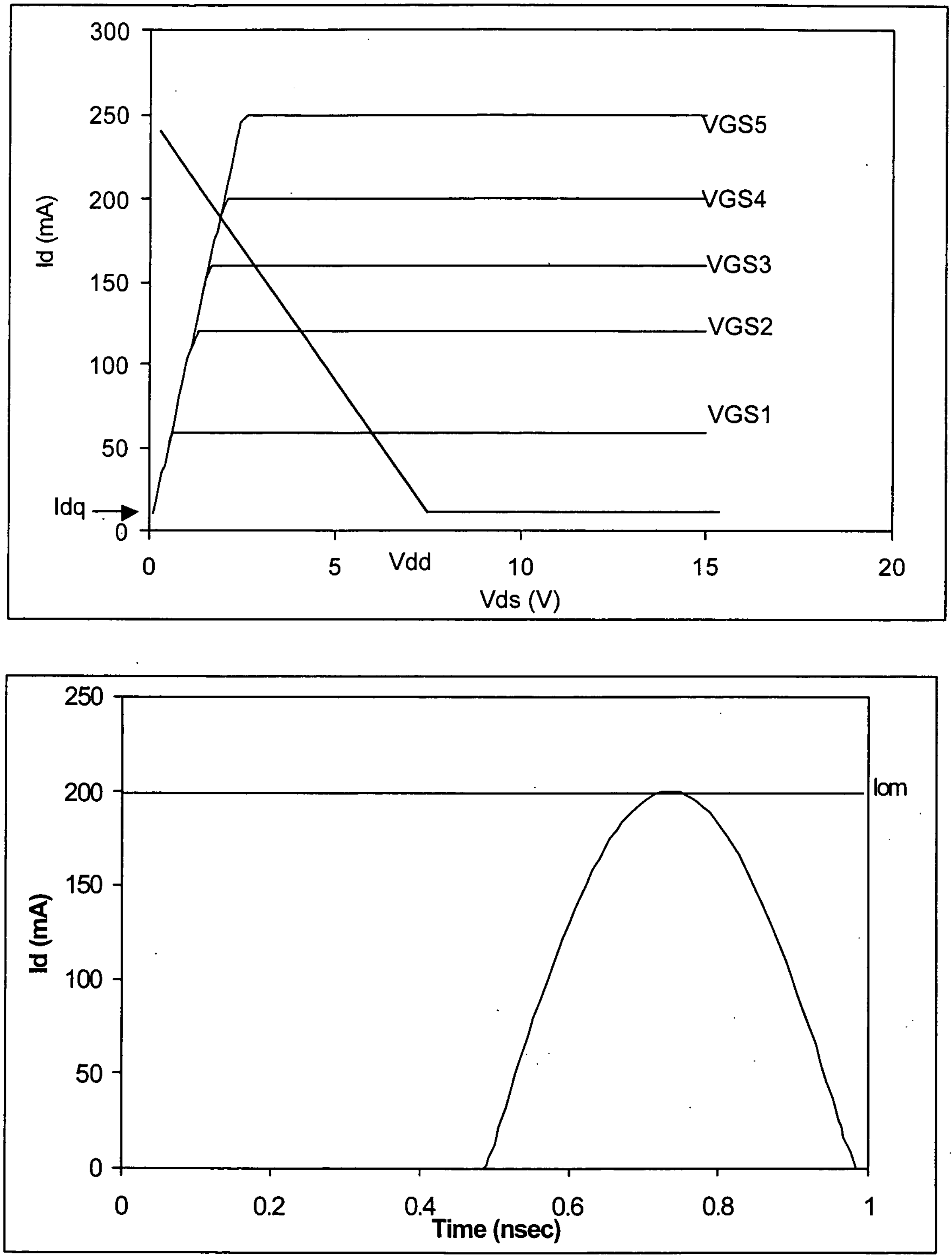


Figure.2.4. **Load line and current waveform for the class-B power amplifier**

2.2.3 Class AB

The class-AB amplifier is a compromise between class A and class B in terms of efficiency and linearity. The transistor is biased as close to pinch-off as possible, typically at 10 to 15 percent of I_{dss}. In this case, the transistor will be on for more than half a cycle, but less than a full cycle of the input signal.

2.2.4 Class C

The previous classes, A, B, and AB are considered linear amplifier, where the output signal's amplitude and phase are linearly related to the input signal's amplitude and phase. In the application where linearity is not an issue, and efficiency is critical, non-linear amplifier classes (C, D, E, or F) are used.

Class-C amplifier is the one biased so that the output current is zero for more than one half of an input sinusoidal signal cycle. Figure 2.5 illustrates the operation of the class-C amplifier. A tuned circuit or filter is a necessary part of the class-C amplifier.

Classes-A, AB, B, and C amplifiers can be defined in terms of the conduction angle *Y* as follows:

$$\text{Class of operation} = \begin{cases} \text{A} \ , \ y = \pi \\ \text{B} \ , \ y = \dfrac{\pi}{2} \\ \text{AB} \ , \ \dfrac{\pi}{2} < y < \pi \\ \text{C} \ , \ y < \dfrac{\pi}{2} \end{cases} \qquad (2.15)$$

The conduction angle is

$$Y = \arccos\left(-\frac{I_{dq}}{I_{dd}}\right) \quad . \tag{2.16}$$

The dc current is

$$I_{dc} = \frac{1}{2\pi}\cdot\int_0^{2\pi} i_D(\theta)d\theta = \frac{1}{\pi}\cdot(I_{dq}\cdot y - I_{dd}\cdot\sin(y)$$

$$= \frac{I_{dd}}{\pi}\cdot(\sin(y) - y\cos(y)) \quad \cdot \tag{2.17}$$

Also, the output voltage (V_O) can be obtained in term of Y as

$$V_o = \frac{1}{2\pi}\cdot\int_0^{2\pi} i_D(\theta)\cdot R\cdot d\theta = \frac{R}{2\pi}\cdot\left[4I_{dq}\cdot\sin(y) + 2I_{dd}\cdot y + I_{dd}\cdot\sin(2y)\right]$$

$$= \frac{I_{dd}\cdot R}{2\pi}\cdot\left[2y - \sin(2y)\right] \quad . \tag{2.18}$$

The output power is

$$P_o = \frac{V_o^2}{R} \quad , \tag{2.19}$$

the dc power is

$$P_{dc} = V_{cc}\cdot I_{dd} \quad , \tag{2.20}$$

and the maximum output voltage V_O is

$$\mathrm{V_{OMAX}} = \mathrm{V_{DD}} \quad \cdot \tag{2.21}$$

From the above equations the maximum efficiency is

$$\eta_{\max} = \frac{P_{OMAX}}{P_i} = \frac{2y - \sin(2y)}{4\cdot\left[\sin(y) - y\cdot\cos(y)\right]} \quad . \tag{2.22}$$

Since the peak drain voltage and drain current are

$$V_{DMAX} = 2V_{DD} \quad , \tag{2.23}$$

and

$$I_{DMAX} = I_{dq} + I_{dd} \tag{2.24}$$

respectively, the power output capability factor is

$$\eta_{\max} = \frac{P_{OMAX}}{V_{DMAX} \cdot I_{DMAX}} = \frac{2y - \sin(2y)}{8\pi \cdot [1 - \cos(y)]} \quad . \tag{2.25}$$

Figure.2.6 shows the maximum efficiency versus the conduction angle. Although it is shown that 100% efficiency is possible, it is impractical because the output power is zero, as shown in Fig.2.7.

Although the preceding analysis was for the single-ended amplifier configuration, a similar analysis can be done for the push-pull amplifier configuration. During the positive half of the signal swing, one device will push the current to the load, and during the negative half signal swing, the other device will pull the current from the load. For example, in a class-B push-pull power amplifier, every device is on for one half of the input cycle, which means that the conduction angle is equal to 180 degrees for each device. This is similar to two class-B single-ended power amplifiers connected in a parallel line. From this observation, it is possible to conclude that the efficiency of the push-pull power amplifier is the same as that of the single-ended power amplifier with the same conduction angle, and the output power capability of the push-pull power amplifier is twice that of the single-ended power amplifier. And this result is due to using two FETs.

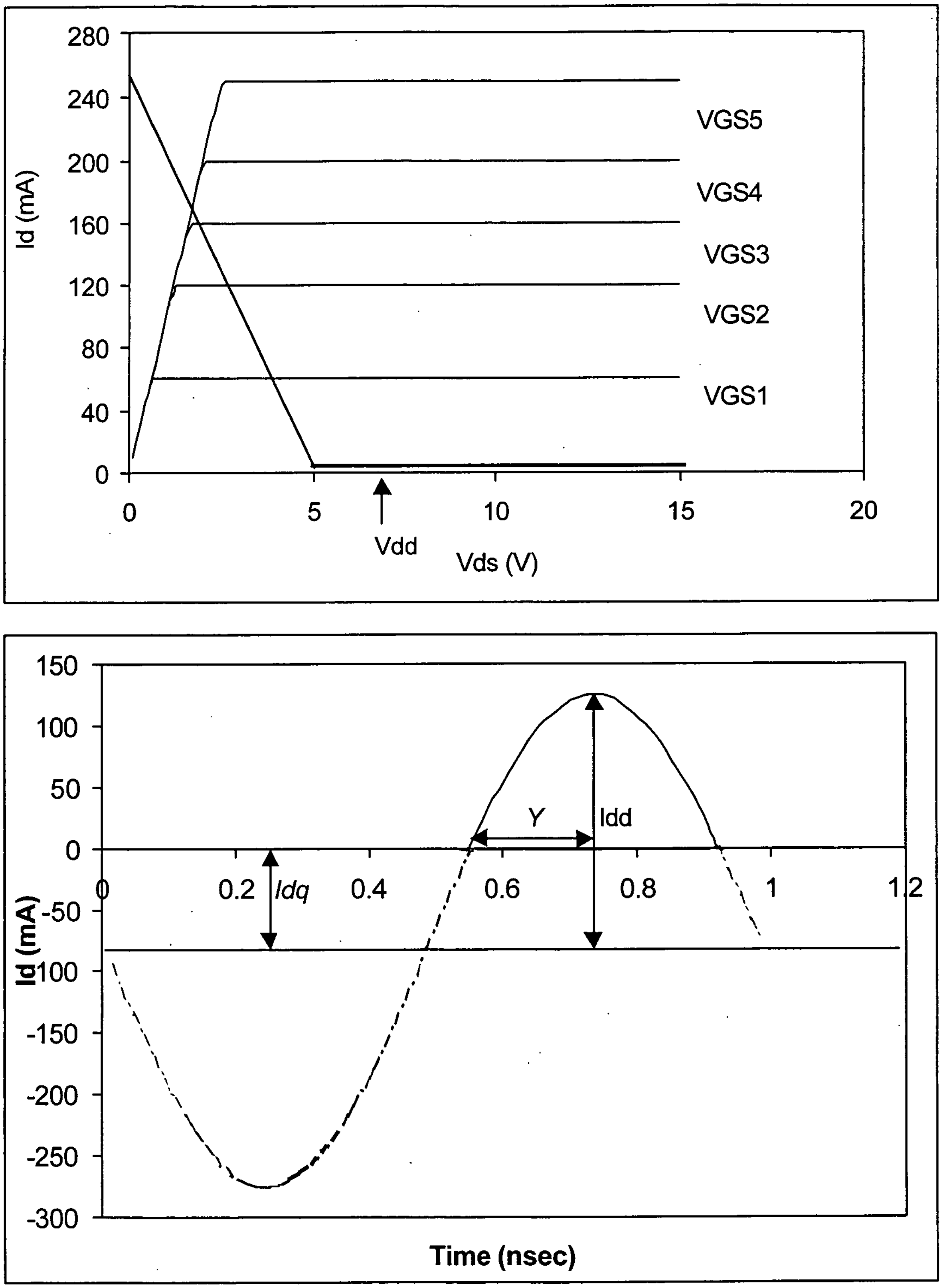


Figure 2.5. **Load line and current waveform for the class-C power**

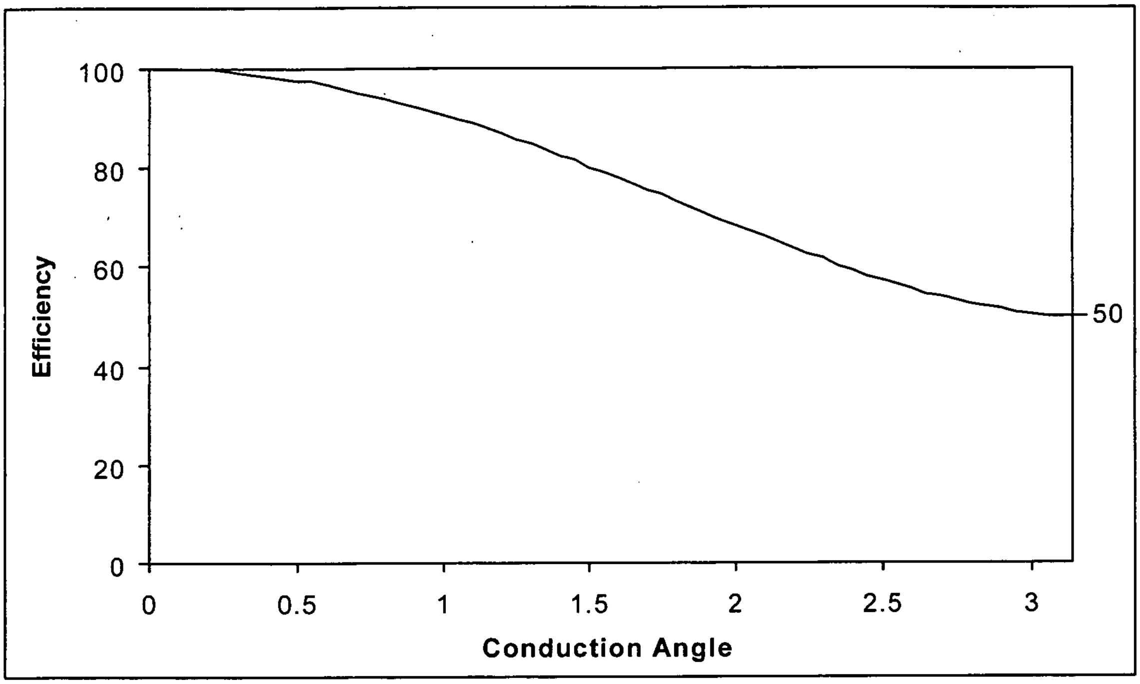


Figure 2.6. **Efficiency vs. conduction angle**

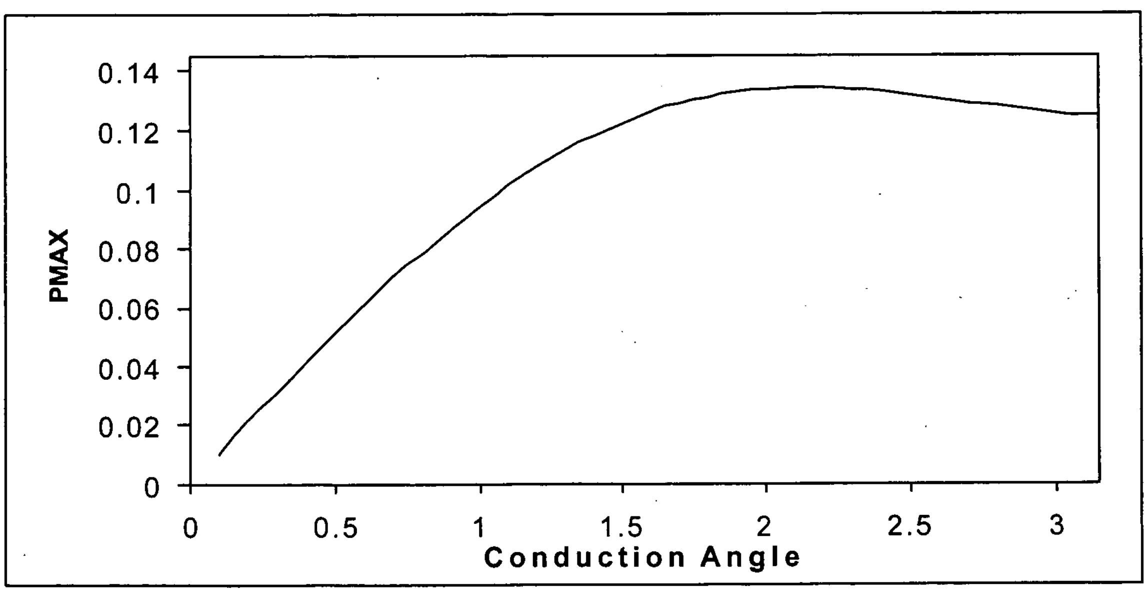


Figure 2.7. **P_{MAX} vs. conduction angle**

2.2.4 Class F

The class-F amplifier is one of the highest efficiency amplifiers. It uses harmonic resonators to achieve high efficiency, which resulted from a low dc voltage current product. In other words, the drain voltage and current are shaped to minimize their overlap region. Figure.2.8 shows a class-F amplifier. The inductor L_3 and capacitor C_3 are used to implement a third harmonic resonator that makes it possible to have a third harmonic component in the collector voltage. The output resonator is used to filter out the harmonic, keeping only the fundamental frequency at the output. The magnitude and the phase of the third harmonic control the flatness of the collector voltage and the power of amplifier.

The drain voltage is

$$V_d(\theta) = V_{DD} + V_{om} \cdot \sin\theta + V_{om3} \cdot \sin(3\theta) \cdot \qquad (2.26)$$

The setting $V_{om3} = \frac{V_{om}}{9}$ produces maximum flatness for the drain voltage. And, the maximum output occurs when the minimum point of $V_d(\theta)$ is zero. Hence,

$$V_{om} = \frac{9}{8} \cdot V_{DD} \cdot \qquad (2.27)$$

The dc current is

$$I_{dc} = \frac{I_{dm}}{\pi}, \qquad (2.28)$$

the dc power is

$$P_{dc} = V_{DD} \cdot \frac{I_{dm}}{\pi}, \qquad (2.29)$$

the fundamental current is

$$I_{om} = \frac{I_{dm}}{2} \cdot \sin\theta \quad , \tag{2.30}$$

the maximum fundamental output power is

$$P_{o\max} = \frac{I_{dm}}{4} \cdot V_{om} \quad , \tag{2.31}$$

and, the maximum efficiency is

$$\eta_{\max} = \frac{P_{o\max}}{P_{dc}} \cdot 100 = \frac{\frac{I_{dm}}{4}}{\frac{I_{dm}}{\pi}} \cdot \frac{\frac{9}{8}V_{DD}}{V_{DD}} \cdot 100 = 88.36\% \quad . \tag{2.32}$$

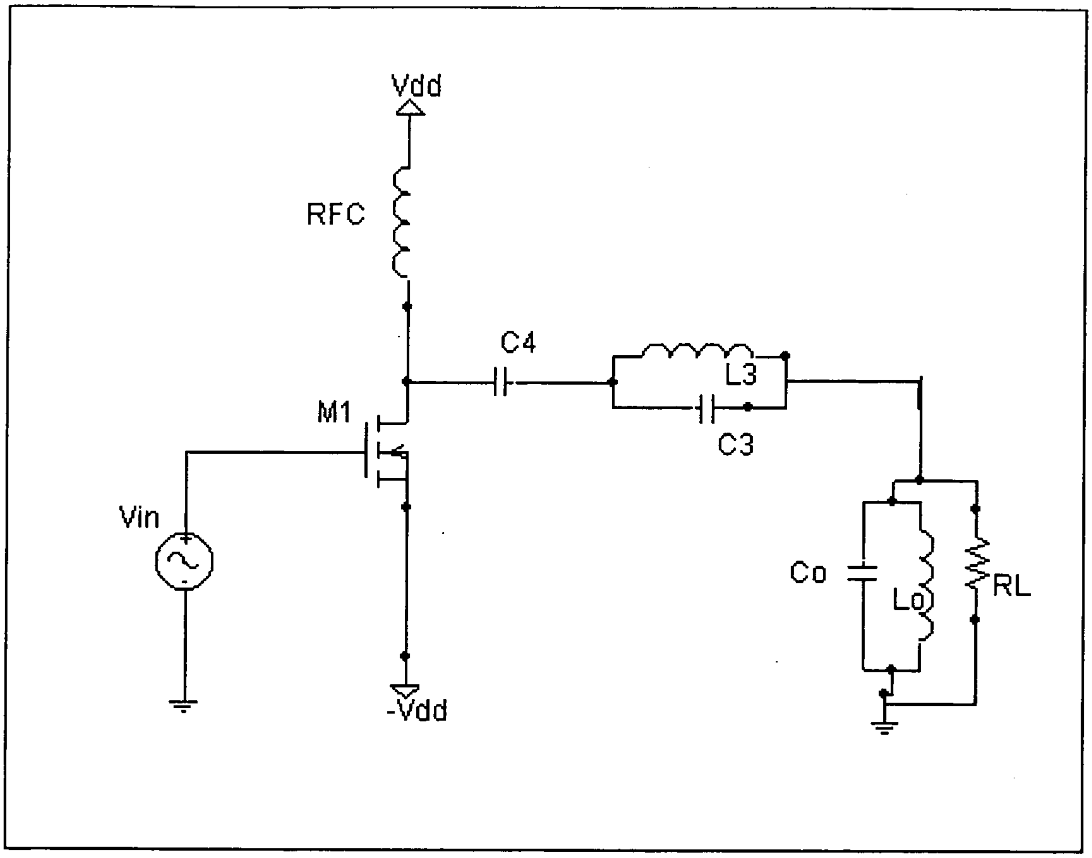


Figure 2.8. **Single-ended power amplifier (class-F)**

2.2.5 Other High-Efficiency Classes

There are other high-efficiency amplifiers such as D, E, G, H, and S. These classes use different techniques to reduce the average collector or drain power, which, in sequence, increase the efficiency. Classes D, E, and S use a switching technique, while classes G and H use resonators and multiple power-supply voltage to reduce the collector current-voltage product. A detailed analysis of class-E amplifier will be presented in Chapter 5.

Designers select the class type to be used based on the application requirements. Classes-A, AB, and B amplifiers have been used for linear applications such as amplitude modulation (AM), single-sideband modulation (SSB), and quadrate amplitude modulation (QAM). Also it can be used in linear and wide band applications such as the multi-carrier power amplifier. Classes C, D, E, F, G, and H have satisfied the need for narrowband tuned amplifiers of higher efficiency. Such applications include amplification of FM signals.

2.3 Main Physical Limitations

The descriptions of amplifiers in the previous sections have dealt with ideal devices. In reality, transistor amplifiers suffer from a number of limitations that influence amplifier operation and ultimately reduce their efficiency and output power.

In practical FET, there are four fundamental effects that force the operation of FET to deviate from the ideal case: the drain source resistance, the maximum channel current I_f, the open channel avalanche breakdown voltage, and the drain-source break down voltage [Robert, 1988]. Figure 2.9 shows I_{DS}-V_{DS} characteristics of a typical MESFET (ATF-46100).

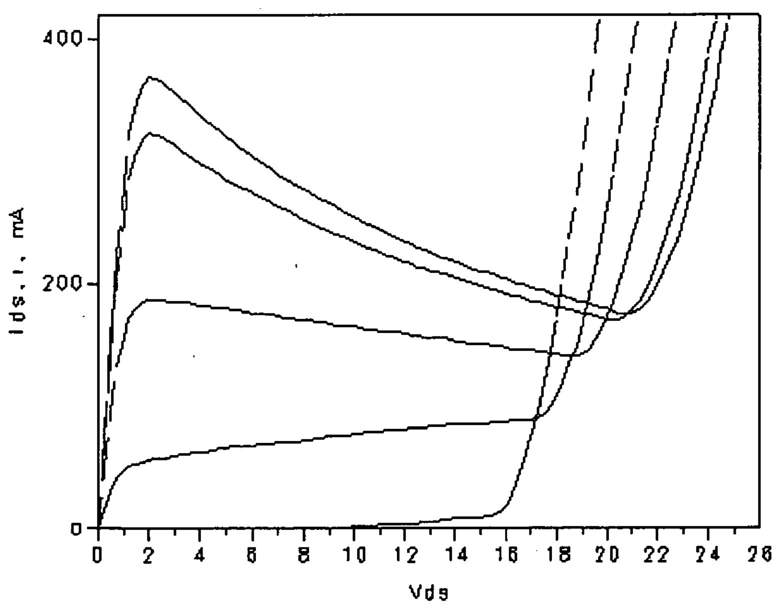


Figure 2.9. **I_{DS}-V_{DS} characteristics of a typical MESFET.**

2.4 Nonlinear MESFET Model

The development of a large-signal model for the transistor is an important step in the design of a nonlinear amplifier. The transistor model consists of linear and nonlinear circuit elements, where the latter are described by a set of nonlinear equations. Figure 2.10 shows a lumped-element model of the MESFET that can be used either in a small-signal or a large-signal analysis. R_g is the ohmic resistance of the gate, and R_s and R_d are the source and drain ohmic resistances, respectively. C_{ds}, and Cg and C_d are the drain-source capacitance and gate-channel capacitances respectively. Several authors have proposed number of nonlinear MESFET models such as the Curtice-Ettenberg model, the Staz model, and the Tom model.

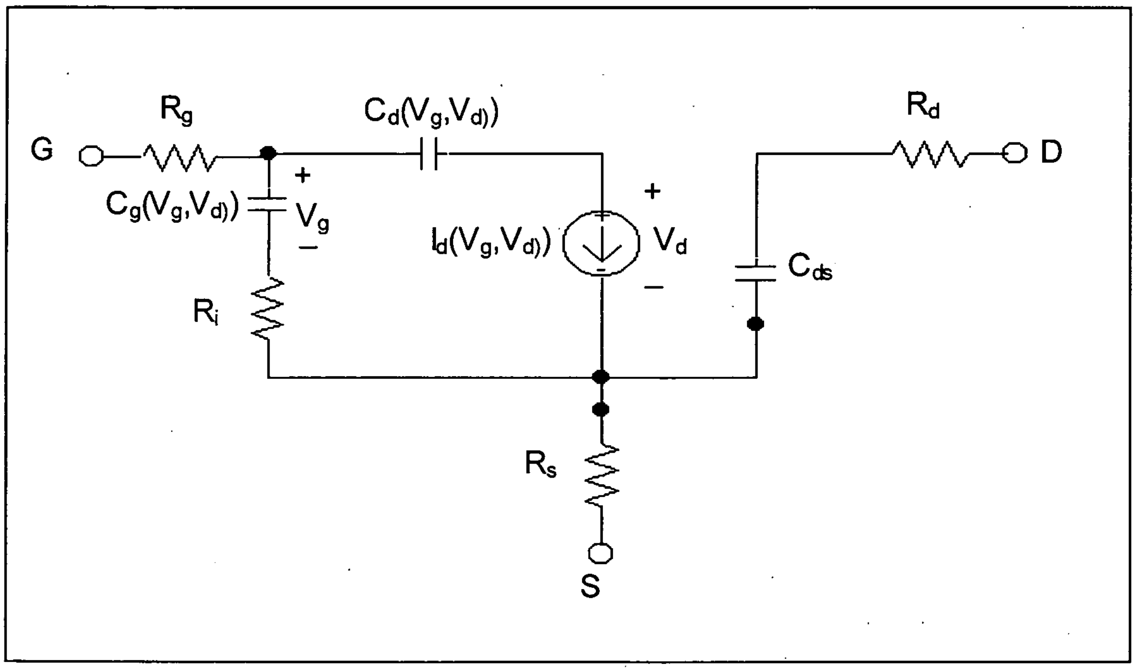


Figure 2.10 GaAs MESFET nonlinear equivalent circuit.

2.5 Nonlinear analysis

Having obtained a suitable model for the large signal behavior of the FET, the next step is to select the analysis method. There are two methods: frequency-domain techniques, and time domain techniques. Harmonic-balance analysis and Volterra-series analysis are the most important frequency-domain techniques. In the harmonic balance technique, the nonlinear circuit is partitioned into linear and nonlinear subcircuits. The linear subcircuit can be described by its Y, S, or any other parameters. The nonlinear elements are modeled by their *I/V* characteristics. The voltages at the interconnections between the two subcircuits are variables that, when determined, define all the voltages and currents in the network. In the Volterra-series analysis, the nonlinear elements are characterized by power series. Then the nonlinear transfer function can be obtained using the convolution.

In the time-domain techniques, conventional circuit theory is used to obtain time-domain differential equations that describe a nonlinear circuit. These differential equations are solved numerically. A major disadvantage of time-domain analysis is that a steady state solution, which is the only one of interest in amplifier design, often requires several cycles and, consequently, a long computation time.

More information about the nonlinear analysis methods for the amplifier circuits is available in Nonlinear Microwave Circuits by Stephen Maas [1987].

Appendix C -- Figures

There are no Figures additional to those in McCallister et al., U.S. Patent No. 6,366,619, relevant to the discussions herein.

Appendix D -- Related Proceedings

Appellant is aware of no related proceedings relevant to this matter.